Differential Amplifiers in Bioimpedance Measurement Systems: A Comparison Based on CMRR

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***Abstract*— In this paper we have analysed the Common Mode Rejection Ratio (CMRR) for differential amplifiers used in bioimpedance measurement systems and derived the complete equations for the case when OPAMPs have finite differential and common mode gains. In principle, passive ac-coupling networks that include no grounded components have an infinite CMRR, but they must provide a path for input bias currents. The paper provides a novel approach as to how component tolerances limit the CMRR and affect the transient response of different networks. Experimental results and various measurements support our theoretical predictions. The best CMRR is obtained when the differential gain is concentrated in the input stage, but it decreases at frequencies above 1 kHz because of the reduced CMRR for the differential stage at these frequencies.**

***Keywords*— differential amplifiers, passive filters, ac-coupling, CMRR, biopotential amplifiers.**

I. Introduction

External bioimpedance measurement on the body uses sinusoidal carrier signals from 10 kHz to 100 kHz. We study the systems that work at a single frequency. The parameter measured is tissue impedance, which is usually in a zone or part where it varies with time. The amplitude of the signal injected is limited for safety regulations to 1 mA at 10 kHz or up to 10 mA at 100 kHz. Typical values of impedance obtained are very small, and, therefore, the need arises for the amplification of the signal obtained before any further processing. OPAMP based differential amplifiers are a common building block in bioimpedance measuring systems. At higher frequencies, discrete transistors usually replace OPAMPS. Differential amplifiers are valuable because of their ability to reject power-line and other common mode interference which follows from their high common mode rejection ratio (CMRR). In this paper, we analysed the CMRR performance of an amplifier with little variations in a basic differential amplifier circuit. First, a theoretical analysis shows what parameters determine the CMRR then we present experimental results that verify the theoretical models developed. We have paid particular attention to phase measurements as they have emerged as being very important in determining whether or not we are obtaining the best CMRR in a given circuit. The tradeoffs between component tolerances, CMRR and transient responses are identified and assessed by experimental measurements.

II. Common Mode Rejection Ratio

The circuit model for 4 terminal bioimpedance measurement systems when using a single ended current source in Fig.1 is analysed to explain the need for high CMRR.

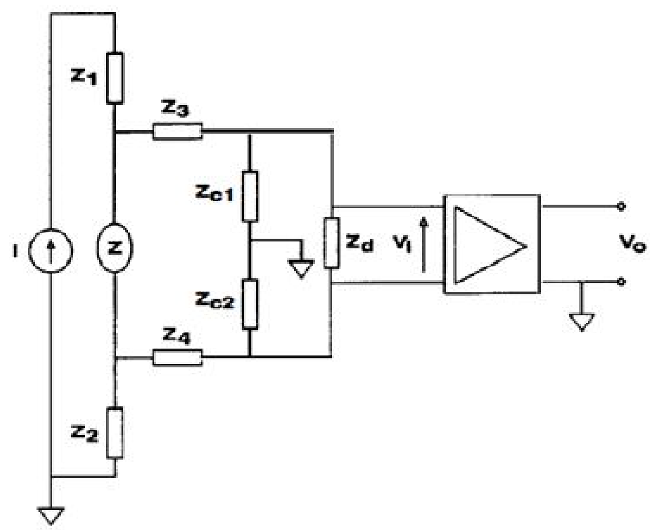

Figure 1. Circuit model for four terminal bioimpedance measurements using single ended current source

A. Infinite Differential and Common Mode Input Impedances

Assuming both the differential and common mode input impedances to be infinite, the amplifier output voltages

$$Vo(s) = H_D(s)I(s)\left\{Z(s)+\frac{1}{CMRR_A(s)}\left[\frac{Z(s)}{2}+Z_2(s)\right]\right\} \quad (1)$$

Here, $H_D(s)$ is the amplifier differential transfer function, $CMRR_A(s)$ is common mode rejection ratio and $Z_2(s)$ is the current injection electrode impedance. Equation (1) shows that only a perfect differential amplifier would result in a zero offset and gain error. In order to evaluate the minimal $CMRR_A$ required in fig (1), we should know Z_2 and its variations. According to [2] we can assume Z_2 to be from 6k&! ||10 nF at 10 kHz to 600&! || 2 nF at 100 kHz, for thoracic impedance plethysmography. To have a 1% offset error, $CMRR_A$ must be atleast about 70dB.

B. Finite Differential and Common Mode Input Impedances

The next step is to determine the system common mode rejection considering finite input impedances. It is well known that in differential systems the total CMRR reduces to a value lower than the CMRR for the amplifier without any input

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imbalances. In bioimpedance measurements, the situation is different from that of biopotential amplifiers as described in [3]: firstly the common and differential mode voltages here are due to the same current (the one being injected), whereas when amplifying biopotentials the most concern is about power line interference; secondly, the ratios between electrode and amplifier input impedances are different from those encountered at low frequencies.

Taking into account the imbalances in electrodes and common mode impedances we define

$Z_3 = Z_e +$ “$Z_e/2$ (2a)

$Z_4 = Z_e -$ “$Z_e/2$ (2b)

$Z_{c1} = Z_c +$ “$Z_c/2$ (3a)

$Z_{c2} = Z_c -$ “$Z_c/2$ (3b)

Where the subindex “e” stands for electrode and “c” stands for common mode. Using the results from [3], we finally have

$$\frac{1}{CMRR_T} \approx \frac{1}{CMRR_A} + \frac{Z_e}{Z_c}\left\{\frac{\Delta Z_c}{Z_c} + \frac{\Delta Z_e}{Z_e}\right\} \tag{4}$$

The common mode input voltage is determined here mainly by Z_2. Therefore, the reduction of Z_2 is an important problem. Feeding back the common mode voltage to the current source (by driving its floating ground) or to the body so that no net current flows through Z_2 is a solution. If the bandwidth is not restricted to about 10 kHz it can result in oscillations. From (4) we deduce that $CMRR_T$ is a complex number. The minimal value of $CMRR_T$ in order to achieve low errors must be judged by considering its phase angle. At the measurement frequencies, both electrode and common mode input impedances have large capacitive components. As a result the phase for $CMRR_T$ will strongly depend on the relative value of impedance imbalances: the lower these are the closer the phase of $CMRR_T$ will be zero. Equation (4) shows that the effect of a given electrode imbalance is smaller for higher common mode input impedances. Therefore, this is another important parameter in amplifier design.

III. Fully Differential Passive AC-Coupling Networks

Fully differential ac amplifiers can easily be designed by placing an ac-coupling network in front of a fully differential dc amplifier [1]. Such a network, however, must have a high common mode rejection ratio (CMRR) to keep the overall CMRR high.

A. *Coupled Single-Ended Filters (Network 1)*

Fig. 1 shows a fully differential ac-coupling network built by joining two single-ended, high-pass filters and connecting the common node to the ground through a high-value resistor R_B [8]. Ideally, R_B should be infinite to achieve an infinite CMRR, but its maximal value is dictated by the input bias currents of the amplifier.

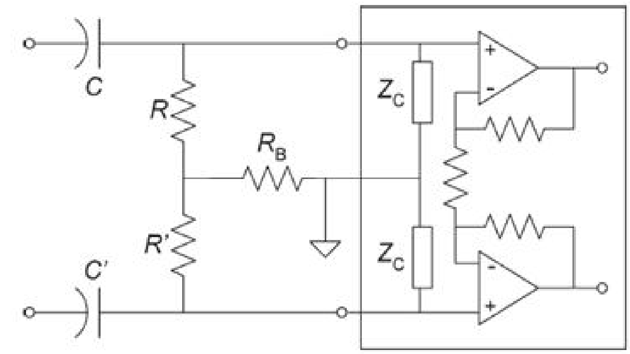

Figure 2. Fully differential passive ac-coupling network built by connecting two single-ended, high-pass filters whose common node is grounded through a high value resistor

If the ensuing common-mode impedances Z_C are large enough, the differential-to-differential gain is given in (1), shown at the bottom of the next page, which for matched components ($R = R\_$, $C = C\_$, and hence, $\tau = RC = R\_C\_ = \tau\_$) simplifies to

$$G_{DD}(s) = \tau s \frac{1 + (\tau + 2\tau_B)s}{[1 + (2\tau_B)s](1 + \tau s)} = \frac{\tau s}{1 + \tau s} \tag{5}$$

where $\tau_B = R_B C$. This transfer function corresponds to a first order high-pass filter. The CMRR for unmatched components when the ensuing common-mode input impedance Z_C is included is

$$CMRR(s) = \frac{(R + 2R_B)CZ_C}{Z_C \frac{\Delta R}{R} + (Z_C + R + 2R_B)\frac{\Delta C}{C} + (R + 2R_B)\frac{\Delta Z_C}{Z_C}} \tag{6}$$

The effect of unmatched components has only been considered in the denominator because its influence on the numerator is irrelevant. If Z_C is mostly determined by the common-mode input impedance C_{IN}, (6) can be rewritten as:

$$CMRR(s) \approx \frac{\frac{\tau + 2\tau_B}{\Delta C/C + \Delta R/R} s}{1 + (\tau_{IN} + 2\tau_{B,IN})\frac{\Delta C/C + \Delta C_{IN}/C_{IN}}{\Delta C/C + \Delta R/R} s} \tag{7}$$

where $\tau_{IN} = RC_{IN}$, and $\tau_{B,IN} = R_B C_{IN}$. At frequencies well below f_c H” $1/[2\pi(R + 2R_B)C_{IN}]$, the CMRR increases by 20 dB/dec according to

$$CMRR_{LF}(s) \approx \frac{(\tau + 2\tau_B)S}{\Delta R/R + \Delta C/C} \tag{8}$$

whereas at frequencies well above f_c, the CMRR reaches the limit value

$$CMRR_{HF}(s) \approx \frac{C/C_{IN}}{\Delta C/C + \Delta C_{IN}/C_{IN}} \tag{9}$$

B. *Cascade-Reversed Single-Ended Filters (Network 2)*

Fig. 2 shows a fully differential ac-coupling network built by cascade-connecting two single-ended high-pass filters with reverse polarity [6, p. 376]. The low-pass version of this circuit has been used for some time in analog front ends for data acquisition [7].

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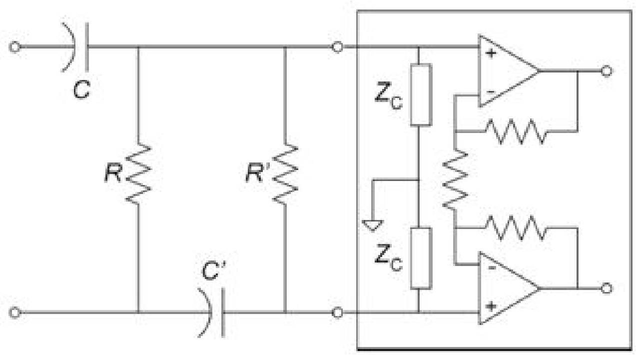

Figure 3. Fully differential passive ac-coupling network built by cascading two single-ended, high-pass filters with reverse polarity

If the ensuing common-mode impedances *ZC* are large enough, the differential-to-differential gain corresponds to a second-order high-pass filter. Hence, the low-frequency rejection is better than that of Network 1. Since, there is no current path to the ground (bias currents flow to the ground through the signal source), the CMRR for the ac-coupling network alone is infinite, even for unmatched components. If finite input capacitances from each amplifier input to the ground are considered, the CMRR is

$$CMRR(s) = \frac{\tau(1 + C/C_{IN})s + 1/2}{\tau(\Delta C_{IN}/C_{IN} + \Delta C/C)s + 1} \quad (10)$$

whose extreme values at very low and very high frequencies, respectively, are

$$CMRR_{LF}(s) \approx \tau(1 + C/C_{IN})s + 1/2 \quad (11)$$

which is limited even for perfectly matched capacitances, and

$$CMRR_{HF}(s) \approx \frac{1 + C/C_{IN}}{\Delta C/C + \Delta C_{IN}/C_{IN}} \quad (12)$$

which is quite close to that for network 1.

C. Coupled Single-Ended Filters Connected to the Input Common-Mode Voltage (Network 3)

Fig. 4 shows a fully differential ac-coupling network built by joining two single-ended, high-pass filters whose common node is connected to the input common-mode voltage obtained by an ungrounded passive voltage adder [8]. If the ensuing common mode impedances Z_c are large enough the differential to differential gain corresponds to a first order high pass filter and no matching is required for R_1 and R_1' to obtain high frequency response. The CMRR for this network would be infinite if the input common-mode impedances of the ensuing amplifier were ideal. If Z_c is mainly determined by stray capacitances C_{IN} then we have the following expression for CMRR.

$$CMRR(s) \approx \frac{C}{C_{IN}} \times \frac{1 + 2\tau s}{\Delta R/R + \Delta C_{IN}/C_{IN} + 2\tau(\Delta C/C + \Delta C_{IN}/C_{IN})s} \quad (13)$$

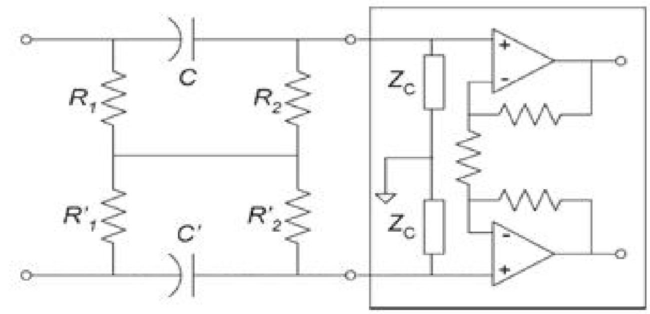

Figure 4. Fully differential passive ac-coupling network built by mirroring two single-ended, high-pass filters whose common node is connected to the common-mode input voltage obtained through an ungrounded passive voltage adder

Whose extreme values at very low and very high frequencies, respectively are

$$CMRR_{LF}(s) \approx \frac{C/C_{IN}}{\Delta R/R + \Delta C_{IN}/C_{IN}} \quad (14)$$

$$CMRR_{HF}(s) \approx \frac{C/C_{IN}}{\Delta C/C + \Delta C_{IN}/C_{IN}} \quad (15)$$

$CMRR_{LF}$ is larger than that for the networks 1 and 2, whereas $CMRR_{HF}$ is uncapped and can become very large for matched capacitances.

D. Another Proposed Active dc Suppression Amplifier Circuit without Grounded Resistor

The circuit shown in fig 5. uses an integrator in a feedback loop around the difference amplifier [1], [6]. The system has two ac-coupled stages: the front differential ac-coupling network and the high pass difference amplifier. The overall transfer function is

$$T(s) = \frac{s\tau_2}{1 + s\tau_2} \frac{s\tau_i AV_0}{1 + s\tau_i} \quad (16)$$

Where $\tau_i = R_i C_i$ and AV0 $= 1 + 2R_4/R_3$. The first factor in (16) corresponds to the passive ac-coupling network and the second factor corresponds to the amplifier and dc restoration circuits.

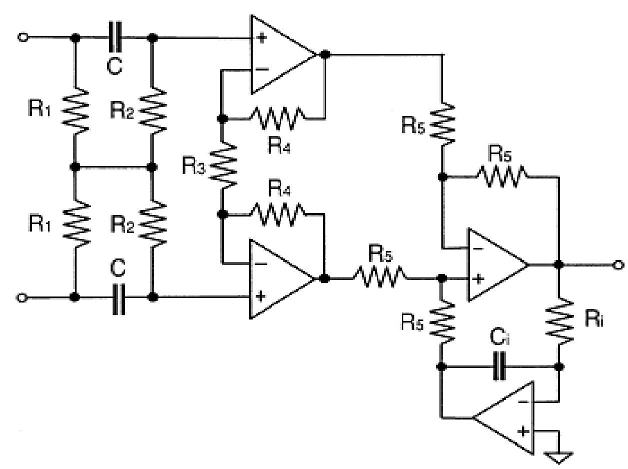

Figure 5. Proposed amplifier circuit, which has two ac-coupling stages, a passive input stage, and an active dc suppression circuit

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The simple, novel, ac-coupled front end for biopotential measurements in Fig. 5 is a fully differential passive-coupling network that does not include any grounded resistor, hence, resulting in a high CMRR. The proposed network enables the design of a high gain for the input stage of biopotential measurement systems, thus leading to implementations with a reduced number of stages, which are particularly convenient for low power applications. Because the common-mode voltage is dc coupled, the proposed circuit also suits single-supply operation.

A single-supply ECG amplifier with a gain of 1001, built according to the design rules proposed and tested for transient and frequency response, and CMRR, fulfilled the requirements in [10], including a CMRR of 123 dB at 50 Hz.

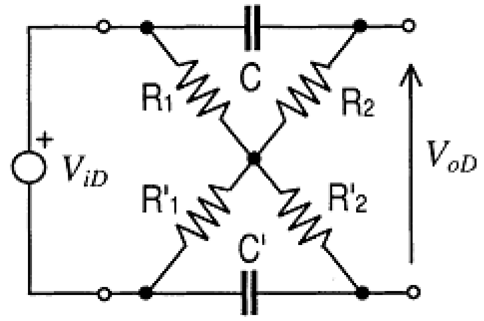

Figure 6. Proposed fully differential ac-coupling network that does not include any grounded resistor

The fig. 6 is an alternative drawing of the proposed passive ac-coupling network in fig. Because it is a fully differential circuit, differential and common-mode voltages define four transfer functions. The main transfer function G_{DD} is the quotient between the Laplace transforms of the differential output voltage V_{OD} and the differential input voltage V_{iD}. The circuit gain is given by the following expression [8]

$$G_{DD} = \frac{1}{s + 1/\tau_2} \qquad (17)$$

where $\tau_2 = R_2C$.

E. Another Circuit with dc Output Voltage Feedback and no Grounded Component

Fig. 7 shows a simplified model of the proposed biopotential amplifier [4]. It includes a fully differential feedback network with two attenuators, one located at the input of the integrator stage and the other at its output.

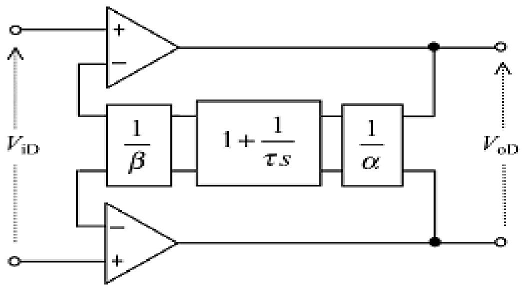

Figure 7. Fully differential circuit for dc suppression implemented by feeding back the dc output voltage using a fully differential network

Assuming a negligible error voltage at the op amps inputs (this implies infinite gain op-amps), the ratio between differential-mode input and output voltages (=GDD, differentia input to differential output gain) is

$$A(s) = \frac{\alpha\beta s\tau}{1 + s\tau} \qquad (18)$$

Fig 8. is a possible implementation of the circuit shown above. Assuming $R_1=R_1'$, $R_2=R_2'$, $R_4=R_4'$, $R_T=R_T'$ and $C_T=C_T'$, we have

$$\alpha = 1+2R_4/R_3 \qquad (19a)$$

$$\beta = 1+2R_2/R_1 \qquad (19b)$$

$$\tau = R_TC_T \qquad (19c)$$

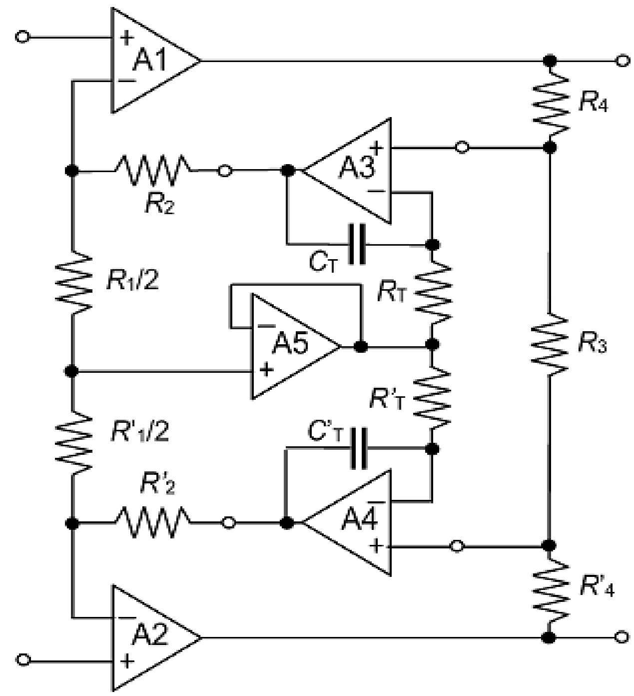

Figure 8. Implementation of the method in Fig. 7. Because the proposed topology does not include any grounded component, its CMRR is ideally infinite

The passive component mismatches do not degrade the CMRR. Given that the proposed circuit has no connections to ground, and disregarding stray capacitances, no current flows when applying a common-mode input voltage. In fact, all circuit nodes reach this potential regardless of eventual mismatches in passive components due to their tolerance. Hence, the differential output is zero, which means $G_{DC}=0$ (common mode input to differential output gain), and the CMRR ($=G_{DD}/G_{DC}$) is infinite. In practice, however, mismatches in differential and common-mode op amp open loop gain limit the global CMRR [5]. A good choice to minimise opamp mismatches is to use dual IC models which yields typical CMRR values greater than 100 dB at 50 Hz even for general purpose opamps.

IV. EXPERIMENTAL RESULTS AND DISCUSSION

The first three passive ac-coupling networks in Figs. 2-4 have been built and tested for the following component

values: R = R' = 1 MΩ , RB = 10 MΩ , C = C' = 100 nF (Network 1);
R = R' = 2.8 MΩ , C = C' = 100 nF (Network 2) ; and $R1$ = $R2$ = 1 MΩ , C = C' = 100 nF (Network 3). These values were selected to set the same high-pass corner frequency of about 1.6 Hz for the three networks. Resistors and capacitors had a ±5% tolerance. The instrumentation amplifier (IA) connected to the network output was AD620AN, whose relevant parameters had the following typical values: IBIAS = 0.5 nA, Ios = 0.3, Vos = 30 μV, CIN = 2 pF, and CMRR = 135 dB at dc for G = 1000. The maximal CMRR that can be measured with our setup is limited by the CMRR of this IA. Therefore, we have experimentally characterized this CMRR from dc to 1 kHz and verified that it is greater than 115 dB. To keep CIN small, IC pins were directly soldered to input components so that no printed circuit board lands would be involved in the connection. Fig. 9 shows the CMRR for the three networks analyzed.

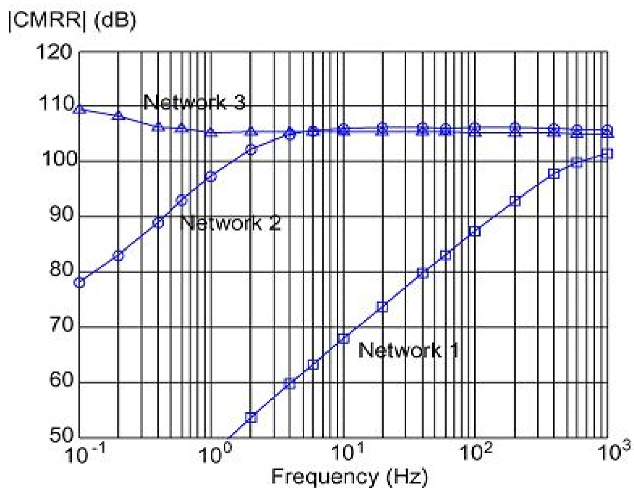

Fig. 9. CMRR for the three ac-coupling networks in Figs. 2-4

The following table summarises a comparative behaviour of the CMRR at low and high frequencies as well as the offset voltages due to input bias currents for the three networks (1-3).

TABLE I.
LOW AND HIGH FREQUENCY CMRR BEHAVIOR AND OFFSET VOLTAGES DUE TO INPUT BIAS CURRENTS FOR THE NETWORKS (1-3)

	Network 1	Network 2	Network 3
$CMRR_{LF}$	$\dfrac{(\tau+2\tau_B)s}{\frac{\Delta R}{R}+\frac{\Delta C}{C}}$	$\tau\left(1+\frac{C}{C_{IN}}\right)s+\frac{1}{2}$	$\dfrac{C/C_{IN}}{\frac{\Delta R}{R}+\frac{\Delta C_{IN}}{C_{IN}}}$
$CMRR_{HF}$	$\dfrac{C/C_{IN}}{\frac{\Delta C}{C}+\frac{\Delta C_{IN}}{C_{IN}}}$	$\dfrac{1+\frac{C}{C_{IN}}}{\frac{\Delta C}{C}+\frac{\Delta C_{IN}}{C_{IN}}}$	$\dfrac{C/C_{IN}}{\frac{\Delta C}{C}+\frac{\Delta C_{IN}}{C_{IN}}}$
V_{bias}	$\lvert I_{BIAS1}R-I_{BIAS2}R'\rvert$	$\lvert I_{BIAS2}R'\rvert$	$\lvert I_{BIAS1}R_2-I_{BIAS2}R_2'\rvert$

For a worst-case condition, with a ±5% tolerance in passive components and a ±10% tolerance in input-amplifier impedance, the theoretical CMRR values at 50 Hz are 70 dB (Network 1) and 105 dB (Networks 2 and 3). Thus the three proposed networks improve the CMRR value of industry-standard ac-coupling circuits that have separate connections from each signal terminal to the ground. The limited input impedance of network 3 can reduce the CMRR.

V. CONCLUSION

We have analyzed five differential ac coupling networks that do not use separate grounded components for each signal line and, hence, achieve a very high CMRR. None of the networks rely on matched passive components. The network 2 (Fig. 3) and network 3 (Fig. 4) yield a better CMRR than network 1 (Fig. 2). The proposed network in Fig. 5 has a reduced number of stages convenient for low power applications and is also suitable for single-supply operations. Lastly, the proposed DC suppression circuit in Fig. 8 , in addition to ac-coupling, provides a simple method for fast restoring the dc level of biopotentials. Also the absence of any grounded passive component makes the CMRR of the amplifier insensitive to the tolerance of passive components. A circuit implementation with a single 5V power supply yields 102dB CMRR at 50 Hz.

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